

Modulator Based on Electronic Change of Phase Shift in Simple Oscillator

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Abstract

In this study, we propose phase shift keying system using an oscillator based on two all-pass filters. The main idea is based on phase commutation by converted data to two pairs of DC control voltages for controlling both all-pass sections. Therefore, the proposal of converter of data to DC control voltages is solved with precise amplitude automatic gain control circuit. The system operation was verified while employing commercially available devices through PSPICE program at the modulation frequencies 1 MHz and 10 kHz with driving signals 0 V (L) and 5 V (H), respectively. The simulation results confirmed the theoretical presumptions.

1. Introduction

The particular problems of multiphase oscillator solutions have been frequently solved in recent works ([1-9] for example). Several works focus on lower multiples of phase shift than of 90 or 60 degrees, for example 45 degrees [6, 8, 9]. There are also implemented some special approaches to the design of the multiphase oscillators [10-12]. For example, Ozoguz *et al.* [10] presented utilization of functional blocks operating in approximated square-root domain for the construction of multiphase oscillator. Maundy *et al.* [11] used fractional-order all-pass (AP) transfer section for the design of multiphase oscillator. Symbolical transfer function of this section contains fractional exponent α of the Laplace operator s , where fractional character is given by RC approximation of the fractional order capacitor in the structure. This feature allows synthesis of the multiphase oscillators with theoretically arbitrary phase shifts between generated signals and extended range of frequency of oscillations (FO) tuning proportional to reciprocal time constant exponentiated by $1/\alpha$. Unfortunately, unavailability of simple electronic adjusting of the fractional-order capacitor limits real utilization of fractional-order-based circuits in systems for the signal generation and processing. There are many problems suitable for future research that have to be solved. A very interesting work regarding multiphase generation based also on non-standard principle was published by Promee *et al.* [12], where log-domain [13] approach was used to obtain reduced complexity of all-pass section, better dynamics of linear operation and wider tunability of oscillation frequency.

Our proposal utilizes the oscillator with theoretically arbitrary phase shift of two generated signals. Generation of arbitrary phase distance between two output signals has been hardly ever solved in simple second-order oscillator structures.

Some notes about arbitrary phase shift generation were given by Tu *et al.* [6]. However, their solutions still produce integer multiples of the minimal phase shift distance between generated amplitudes. The first attempt to really continuously settable phase distance of two produced signals in the frame of the simple oscillator was presented in [14]. However, no other ideas how to utilize oscillator with such feature were discussed. Our question is: What application/system can be prepared with oscillator producing “arbitrary” phase shift? Therefore, the presented work focuses on simple application of the modified oscillator structure (voltage controlled all-pass sections including precise amplitude automatic gain control circuit) in simple phase shift keying modulator system including control logic (converter of data to DC driving voltages).

2. Proposed Solution

Our aim is to design modulator with two outputs providing identical carrier frequency and with digitally controlled phase shift between them (phase shift commutated between two values). We used modification of the simple oscillator based on two all-pass filtering sections [14] as the main building block of the solution. However, several other counterparts are required for the design. First of all, precise automatic amplitude gain control circuit (AGC) has to be designed in order to ensure and preserve stable output levels. We want to implement oscillator solution with voltage-controlled phase shift to achieve the simplest method of driving. In such a case the data (symbols) are represented by square wave with levels 0 V (logic L) and 5 V (logic H). Therefore, the next important task is the proposal of “converter of data (symbol) to DC control voltage”. Therefore, we divided our work to several subparts where all these partial problems will be discussed separately. The block structure of the system is shown in Fig. 1.

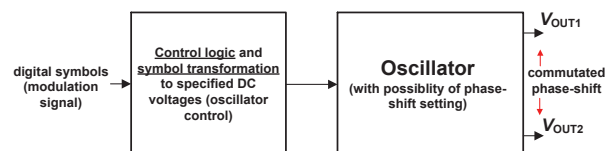


Fig. 1. Proposed concept of digital modulator

2.1. Oscillator with Possibility of Phase Shift Setting

The oscillator structure is shown in Fig. 2. It utilizes a solution based on two AP sections (derived from [15, 16]) and

inverting voltage amplifier presented in [14]. Note that solution in **Fig. 2** has many different features (including controllable amplifier VA and a different type of the phase shifters) than [14].

The condition of oscillation (CO) is fulfilled for $-A \geq 1$. Variable voltage amplifier VCA610 [17], driven by negative DC voltage, serves for control of CO. FO is given by:

$$\omega_0 = \sqrt{\frac{B_1 B_2}{R_1 R_2 C_1 C_2}} \cong \sqrt{\frac{V_{SETB1} V_{SETB2}}{R_1 R_2 C_1 C_2}}. \quad (1)$$

It is clear that FO of the oscillator is electronically adjustable (even linearly if $V_{SETB1} = V_{SETB2} = V_{SETB}$). Phase relations of the operation of this oscillator can be derived from transfer relations of partial AP sections. This construction of AP is based on electronically controllable current conveyor of second generation (ECCII+) [18, 19] built with help of current-mode multiplier EL2082 [20], very popular diamond transistor (DT) OPA660 [21] and differential voltage buffer (DVB) AD830 [22] with unity gain transfer.

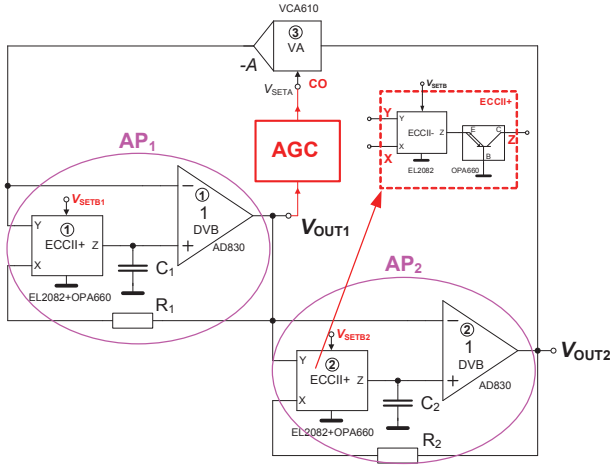


Fig. 2. Oscillator used in proposed solution

By routine analysis we can obtain phase shift between input and output of the AP section (we suppose fulfilled CO, i.e. $-A = 1$):

$$K_{AP1}(s) = \frac{\frac{B_1}{R_1} - sC_1}{\frac{B_1}{R_1} + sC_1} = \frac{\sqrt{\left(\frac{B_1}{R_1}\right)^2 + (\omega_0 C_1)^2} e^{\tan^{-1}\left(\frac{\omega_0 C_1 R_1}{B_1}\right)j}}{\sqrt{\left(\frac{B_1}{R_1}\right)^2 + (\omega_0 C_1)^2} e^{\tan^{-1}\left(\frac{\omega_0 C_1 R_1}{B_1}\right)j}} = 1 \cdot \exp\left[-2 \tan^{-1}\left(\frac{\omega_0 C_1 R_1}{B_1}\right)j\right]. \quad (2)$$

Supposing insertion of ω_0 from (1) to (2), we obtained argument of the equation in form:

$$\varphi_{1-2} = -2 \tan^{-1}\left(\sqrt{\frac{B_2}{B_1} \frac{R_1 C_1}{R_2 C_2}}\right)_{\substack{R_1=R_2=R \\ C_1=C_2=C}} \cong -2 \tan^{-1}\left(\sqrt{\frac{B_2}{B_1}}\right). \quad (3)$$

It can be seen that phase shift of the produced signals depends on B_2/B_1 ratio. Therefore, we can use these two parameters for electronic control of phase shift between V_{OUT1} and V_{OUT2} and for control of FO (if defined ratio is kept constant during the tuning process). Note that also other parameters (R_1 , R_2 especially) may serve for definition of phase distance (ratio) and

then both identical values of $B_{1,2}$ (adjusted simultaneously) may serve for FO control.

We selected the following parameters for validation and demonstration of the assumed features of the oscillator: $C_1 = C_2 = C = 220$ pF, $R_1 = R_2 = R = 565 \Omega$ ($470 \Omega + 95 \Omega$ of the X terminal [20]). Our design requirements are given as follows: $\varphi_{1-2} = 30$ deg in case of logic L (0 V, as explained in section 2.2) at the data input and $f_0 = 1$ MHz. Supposing that (3) can be rewritten to:

$$\frac{B_2}{B_1} = \tan^2\left(-\frac{\varphi_{1-2}}{2}\right), \quad (4)$$

and considering (1), we have two possibilities of driving the commutation:

$$\frac{B_2}{B_1} = \text{const.}, \quad \frac{B_2}{B_1} = \frac{1}{\text{const.}}, \quad (5), (6)$$

while product $B_1 B_2$ is always the same. Therefore, we found out that the second phase distance of V_{OUT1} and V_{OUT2} (for logic H – 5 V at the data input, as explained in section 2.2) is given by mutual dependence of B_1 and B_2 to keep their ratio directly driving FO. We can define only initial phase distance (for L). The second phase distance for H is given by mutual dependence of (1) and (3). Therefore (3) or (4) and (1) does not allow to select the second phase distance, i.e. product of B_1 and B_2 has to be the same in both cases of phase shift commutation to obtain unchangeable FO. The first value of B_1 can be calculated (by substitution of (1) and (4)) as:

$$B_1 = \frac{2\pi f_0 \sqrt{R_1 R_2 C_1 C_2}}{\tan\left(-\frac{\varphi_{1-2}}{2}\right)} \Bigg|_{\substack{R_1=R_2=R \\ C_1=C_2=C}} = \frac{2\pi f_0 RC}{\tan\left(-\frac{\varphi_{1-2}}{2}\right)}. \quad (7)$$

Numerical calculation yields $B_1 = 2.91$ for $\varphi_{1-2} = 30$ deg. Due to nonlinear dependence of B on V_{SETB} [20] above 2 V, V_{SETB1} must be 3.4 V in this case. The second gain/DC voltage can be expressed from (4) as $B_2 = 0.21$ ($V_{SETB2} = 0.21$ V). So, we obtained two sets (pairs) of values $B_1 = 2.91$ ($V_{SETB1} = 3.4$ V), $B_2 = 0.21$ ($V_{SETB2} = 0.21$ V) for L (data) and $B_1 = 0.21$ ($V_{SETB1} = 0.21$ V), $B_2 = 2.91$ ($V_{SETB2} = 3.4$ V) for H from (5) and (6). The phase distance of V_{OUT1} and V_{OUT2} in the H is 150 deg.

2.2. Control Logic – Converter of data symbols to DC control voltage

Due to the requirements on the oscillator and AP blocks of the structure to obtain phase shift commutation, we have to ensure two driving signals for control of the current gains $B_{1,2}$ where two specific values (levels) are available. In other words, our task is to design a controlling logic part or “data to voltage converter” with two outputs, where commutation of two V_{SETB} (0.21 V and 3.4 V) in dependence on the state of input signal (L or H) is possible. Therefore, we designed system shown in **Fig. 3**. The presented solution contains two sections/channels (for V_{SETB1} and V_{SETB2}). The first part of each channel is a comparator part where input modulation signal (data) is compared with reference level 1.5 V (to decide whether input signal is logic 0 or 1, i.e. 0 V or 5 V). In case that input level has character of log. 0, one comparator from the pair has output in high state (saturation – $V_{sat_LM239} = 4.85$ V) and the second in low state (0 V). These results are evaluated by adder/subtractor

based on simple opamp. The amplification/attenuation is set by resistors to ensure the appropriate result of subtraction (0.21 V for L and 3.4 V for H). Therefore, DC voltage of the reference level referencing the both subtraction stages is -4.85 V.

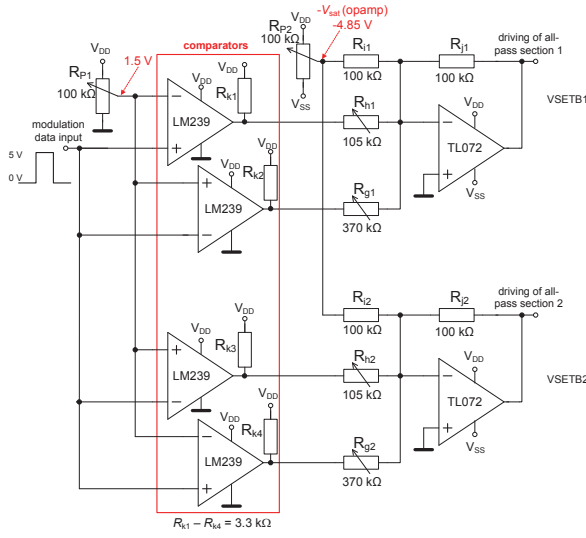


Fig. 3. Designed converter of data symbols to DC control voltage

Based on this, we can calculate values of resistors (resistors $R_{i1} = R_{j1}$ in the direct path with inverting unity gain were selected as 100 kΩ) of the subtractor as:

$$R_{g1} = \frac{R_{j1}}{\left(1 - \frac{V_{SETB1_L}}{V_{sat_LM239}}\right)}, \quad R_{h1} = \frac{R_{j1}}{\left(1 - \frac{V_{SETB1_H}}{V_{sat_LM239}}\right)}, \quad (8), (9)$$

that yields $R_{g1} = 334$ kΩ (for $V_{SETB1_L} = 0.21$ V) and $R_{h1} = 105$ kΩ (for $V_{SETB1_H} = 3.4$ V). Value 334 kΩ was modified to 370 kΩ for correct setting of $V_{SETB1_L} = 0.21$ V in simulations. The second section employs the same values ($R_{h2} = R_{h1}$ and $R_{g2} = R_{g1}$) but their timing is inverted, i.e. while L causes $V_{SETB1} = 0.21$ V, V_{SETB2} has value 3.4 V and in the case of H V_{SETB1} is 3.4 V and $V_{SETB2} = 0.21$ V. Simple modification of R_g , R_h values allows to set up output levels of V_{SETB} quite simply.

2.3. Automatic Amplitude Gain Control Circuit

Practical implementation of each oscillator working in low and middle frequency bands (except GHz band where different principles are used) requires proposal of amplitude automatic gain control circuit (AGC). We implemented simple structure based on the diode doubler (envelope detector) and two simple opamps in **Fig. 4**. DC constant value 1.1 V represents DC voltage for control of CO ($V_{SETA} = -1.1$ V in **Fig. 2**). Output oscillations (at AGC input) are transformed to dynamically (slowly) changing DC voltage signal from rectifier (doubler) and this signal is amplified/attenuated by inverting amplifier and summed with DC constant 1.1 V. Therefore, AGC responses on signal amplitude change by appropriate increase or decrease of the DC control voltage (V_{SETA}) for VA in **Fig. 2**.

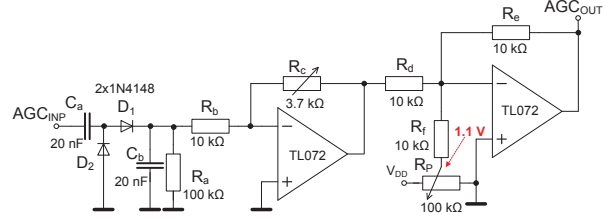


Fig. 4. Automatic amplitude gain control circuit (AGC)

3. Results of Simulations

We verified function of the modulator system in transient responses where digital input data (symbols) with frequency 10 kHz, level 0 V (L) or 5 V (H) and duty cycle 50% are directly driving the phase shift distance of both output signals (V_{OUT1} , V_{OUT2}), **Fig. 5**. It also includes both driving signals from “converter of data to control voltage” used for setting of B_1 (V_{SETB1}) and B_2 (V_{SETB2}). The last result of our simulation is transient response at the both outputs of the oscillator (V_{OUT1} , V_{OUT2}). All these results are shown for FO equal to 1.055 MHz (simulated) in **Fig. 5**. The detail on edge of the driving signal (data) to see the reaction of the oscillator on change of the phase distance is given in **Fig. 6**.

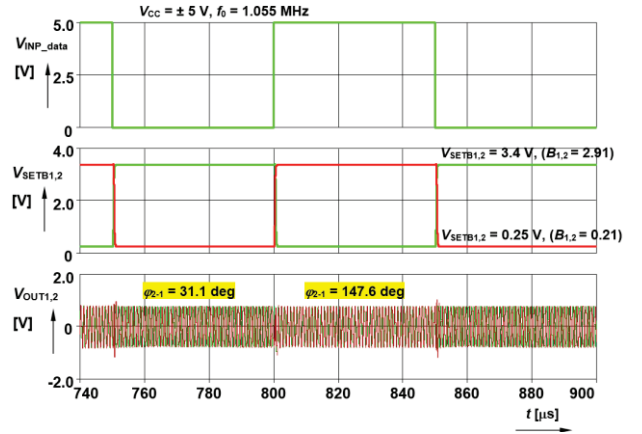


Fig. 5. Transient response of the modulator system on the data input – phase distance commutation

Response seems to be a little bit delayed (0.5 μs) due to the delay of the rising and falling edges of the voltages V_{SETB1} and V_{SETB2} from converter part (simple opamps with low slew rate were used in the converter in this particular solution). Stabilization of the level and waveforms takes about 1.5 μs. FFT spectrum of the output responses (V_{OUT1} , V_{OUT2}) with modulation (**Fig. 6**) is shown in **Fig. 7**. **Figure 8** documents spectrum of the signals without modulation. It is clear and expected that modulation causes worse phase noise features (energy is spread to wider bandwidth), as obvious from the comparison of **Fig. 7** and **Fig. 8**. Spectral purity of the output responses (**Fig. 8**) is very high. Note that values of Y-axis in **Fig. 7** and **Fig. 8** are in dBV ($20\log_{10}(V_{OUT1,2}/1$ V)). Suppression of higher harmonic components is more than 60 dBc that yields total harmonic distortion below 0.1 %.

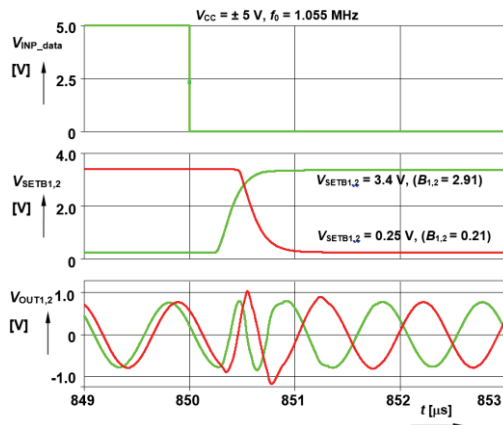


Fig. 6. Detail on the transient effect near to falling edge of the driving signal (data) – phase distance commutation

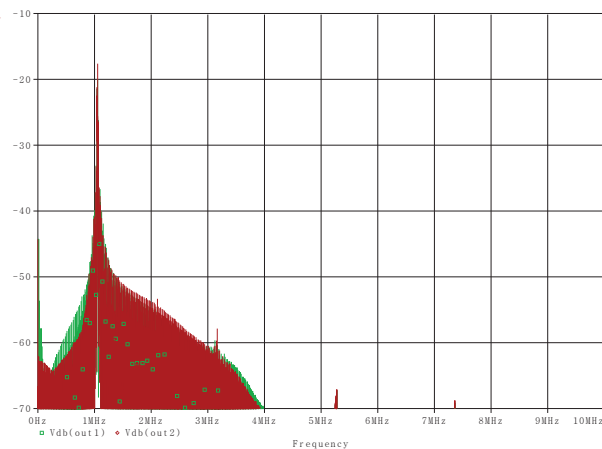


Fig. 7. FFT spectrum of the output signals (VOUT1, VOUT2) under condition of modulation

6. Conclusions

Our work shows how to utilize the interesting oscillator with adjustability of the phase distance between produced signals in the digital phase shift keying modulator. All required construction counterparts were carefully and detailed discussed. Design of the phase commutation was provided for the initial ideal phase shift 30 degrees for input state L and the frequency of oscillation equal to 1 MHz. The second ideal phase shift is 150 degrees in opposite state (assumed for H). The simulations verified and confirmed expected behavior at 1.055 MHz. with two available phase shifts 31.1 and 147.6 degrees. Thus, our verifications confirmed expected behavior of the system. The system has some limitations: the second phase shift (for H) cannot be selected arbitrarily; inaccurate setting (commutation of control levels) may cause parasitic change of FO. These issues are the main aim of our further research and analyses as well as establishment of the experimental verification in order to prove the workability of the design in lab environment.

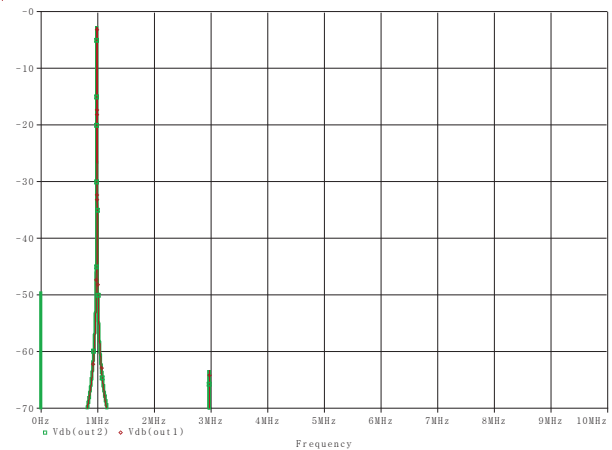


Fig. 8. FFT spectrum of the output signals (VOUT1, VOUT2) without modulation

7. Acknowledgement

Research described in this paper was financed by Czech Ministry of Education in frame of National Sustainability Program under grant LO1401. For research, infrastructure of the SIX Center was used. Research described in the paper was supported by Czech Science Foundation project under No. 14-24186P. Grant No. FEKT-S-14-2281 also supported this research. The support of the project CZ.1.07/2.3.00/20.0007 WICOMT, financed from the operational program Education for competitiveness, is gratefully acknowledged.

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